

*This article is identical to the version that appeared in Appl. Phys. Lett. **88**, 203506 (2006) .*

The published version can be found at <http://apl.aip.org/>.

Radio-frequency point-contact electrometer

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(Dated: February 1, 2008)

Abstract

We fabricate and characterize a radio-frequency semiconductor point-contact electrometer (RF-PC) analogous to radio-frequency single-electron transistors [RF-SETs, see Science **280**, 1238 (1998)]. The point contact is formed by surface Schottky gates in a two-dimensional electron gas (2DEG) in an AlGaAs/GaAs heterostructure. In the present setup, the PC is operating as a simple voltage-controlled resistor rather than a quantum point contact (QPC) and demonstrates a charge-sensitivity about $2 \times 10^{-1} e/\sqrt{\text{Hz}}$ at a bandwidth of 30 kHz without the use of a cryogenic RF preamplifier. Since the impedance of a typical point-contact device is much lower than the impedance of the typical SET, a semiconductor-based RF-PC, equipped with practical cryogenic RF preamplifiers, could realize an ultra-fast and ultra-sensitive electrometer.

PACS numbers: 07.50.Qx, 85.30.Hi, 85.30.Be, 85.35.Gv, 07.50.Ls

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A fast and sensitive electrometer is desirable for sensor applications, for studying charge dynamics in mesoscopic electron devices and for read-out devices in quantum-information processing. The radio-frequency single-electron transistor (RF-SET), invented by Schoelkopf *et al* in 1998, is the most sensitive *and* fastest electrometer known in the literature [1, 2, 3]. The power of such an RF-SET comes from the integration of an aluminum SET, with ultra-high charge sensitivity, into an RF transmission line, with an embedded LC resonant circuit [1]. Similar ideas have been widely adopted in many different applications, e.g., to sense the single-electron tunneling in a single-wall carbon nanotube [4] or to count moving particles in a micro-fluidic channel [5]. So far, RF-SETs are commonly made of aluminum and fabricated close to the devices under test. RF-SET electrometers made of host semiconductor materials will significantly simplify the fabrication procedure and the architecture of integrated device-electrometer system. However, a practical LC-resonant circuit embedded in a $50\ \Omega$ transmission line requires the SET impedance to be low enough ($R < 100\ \text{k}\Omega$) to maintain a high sensitivity (see below). Although there is one successful semiconductor-based RF-SET reported [6], it turns out very challenging to build an RF-SET from resistive semiconductor SETs ($R \sim 500\ \text{k}\Omega$) than to make one from metallic counterparts ($R \sim 50\ \text{k}\Omega$).

Unlike SETs, which require large tunnel barrier impedances in order to localize the electron on the island, a point contact (PC) operates as a conventional field-effect transistor but with very low gate-channel capacitance and very low channel charge. This enables the PC to use a much lower channel resistance than is possible with SETs and so is ideal for integrating into a resonant RF transmission line (forming an RF-PC analogous to RF-SETs). Under appropriate conditions, the conductance of a PC becomes quantized: $G_N = 2e^2 N/h = N \times (12.9\ \text{k}\Omega)^{-1}$, $N = 1, 2, \dots$, and the device is called quantum point contact (QPC). The sharp transition regions between neighboring conductance plateaus allow even greater sensitivity in the detection of charge fluctuations in nearby electron devices to be realized [7, 8]. Recent results reported in Ref. [9] have shown real-time observations of single-electron tunneling events in a quantum dot detected by a capacitively coupled QPC electrometer. In that experiment, a more direct approach was applied: a room temperature $I - V$ converter was applied to monitor the electrical current through the QPC. Here we demonstrate the first implementation of an RF-PC electrometer by integrating a PC device with a resonant RF transmission line. The electrometer was characterized at both dc and

RF limit.

Fig. 1(a) shows the circuit diagram of an RF-PC. Two inductors and a PC device are embedded in the RF transmission line. The inductors (L) and the capacitance (C) of the PC form an LC resonant circuit. The LC circuit introduces a discontinuity in the electromagnetic impedance in the transmission line (with a characteristic impedance z_0), which reflects an incident RF signal. The reflection coefficient is $\Gamma = (z - z_0)/(z + z_0)$, where z is the input impedance: $z = z_1 + z_2(z_0 + z_1)/(z_0 + z_1 + z_2)$ with $z_1 = j2\pi fL$, $z_2 = R_q/(1 + j2\pi fR_qC)$, and $R_q = R_q(V_g)$ being the resistance of the PC, which is controlled by the gate voltage V_g . Neglecting any phase shift, the transmitted signal is given by $v_o = T(f, R_q)v_i \cos(2\pi ft)$, where $v_i \cos(2\pi ft)$ is the input RF signal and $T(f, R_q)$ is the transmission coefficient: $T(f, R_q) = \{(1 - |\Gamma|^2)/[1 + z_0/R_q + (2\pi f)^2 L^2/R_q z_0]\}^{1/2}$. This circuit has a resonance at $f_0 \approx (2/LC)^{1/2}/2\pi$ with a Q -factor of $Q = 2\pi f_0 L/z_0$. Since the transmission coefficient at resonance, $T(f_0, R_q)$, is controlled by the gate voltage V_g , a small ac signal $v_g \cos(2\pi \xi t)$, superimposed on a dc working-point gate voltage V_g^0 , will generate two sidebands (at $f_0 \pm \xi$) in the output spectrum: $v_{os} = \frac{v_i v_g}{4} \left\{ \frac{dT(f_0, R_q)}{dR_q} \left| \frac{dR_q}{dV_g} \right| \right\}_{V_g=V_g^0} \cos[2\pi(f_0 \pm \xi)t]$. The sensitivity of the output signal to small changes in the gate voltage is proportional to both dT/dR_q and $|dR_q/dV_g|$. The first term is determined by the characteristics of the RF transmission line, while the second term is determined by the characteristic of the PC.

In our experiment, the RF transmission line is realized as a coplanar waveguide ($z_0 = 50 \Omega$) on a high-frequency printed-circuit board (PCB), as shown in Fig. 1(b). Two high- Q on-chip inductors ($L = 100$ nH) are inserted into the coplanar waveguide with small spacing. The semiconductor substrate containing the PC device is mounted and electrically connected to the PCB next to the junction of the two inductors, circuit using short gold bond wires. The PC device was fabricated using a two-dimensional electron gas (2DEG) AlGaAs/GaAs heterostructure, with an electron density of $1.7 \times 10^{15} \text{ m}^{-2}$ about 90 nm beneath the surface and a mobility of $80 \text{ m}^2/\text{Vs}$ at 4.2 K. The PC is defined by two reverse-biased Schottky gates connected to the coplanar waveguide, as shown in Fig. 1. The entire circuit is enclosed in an oxygen-free-high-conductivity copper box, with four SMA connectors. The input RF signal was supplied by either an HP 4396A network/spectrum analyzer or an HP 83711A synthesized continuous-wave generator. The input power was fixed at -40 dBm, which corresponds to an ac voltage of $v_i = 3.2$ mV. The transmitted RF signal was monitored by the spectrum analyzer. No net drain-source bias was applied

in this experiment ($V_{ds} = 0$). However, a 17 Hz sinusoidal signal with the root-mean-square amplitude of $158 \mu\text{V}$ was superimposed onto the input RF signal using a bias tee. This allowed of the measurement of PC's differential conductance ($G = dI_{ds}/dV_{ds} \approx 1/R_q$) by an ITHACO 1211 current preamplifier together with an EG&G 5210 lock-in amplifier. All measurements were performed at 4.2 K.

The transmission coefficient as a function of the frequency measured by the network analyzer is shown in Fig. 1(c). The resonance at $f_0 = 810$ MHz suggests a total capacitance of $C \approx 0.786$ pF [10]. The resonance has a Q -factor about 10. The dashed line in Fig. 1(c) is a simulation based on the circuit model shown in Fig. 1(a) given $R_q = 100$ M Ω . In addition to the attenuation in the point contact, there is also an attenuation of about 6 dB in the coaxial cables, the coplanar waveguide and the inductors.

By tuning the dc gate voltage, the PC's resistance and hence the amplitude of the transmission are varied, as shown in Fig. 1(c). A detailed correspondence between the resistance and the resonant transmission coefficient $[T(f_0)]$ is shown in Fig. 2(a). The resistance increases with a more negative gate voltage and finally reaches the pinch-off state at around -1.65 V. Accordingly, the transmission is enhanced by the increase of resistance. However, the measured transmission does not saturate beyond the pinch-off voltage and is lower than the calculated values based on the lumped-circuit model shown in Fig. 1(a). This discrepancy is caused by the assumptions of this simple circuit model: The resistance measured at quasi-dc limit and a shunted capacitance are not sufficient for modeling the RF characteristics. A network of stray capacitors and resistors could be a better model; nevertheless the present model fits very well to the data for $R_q < 23$ k Ω (i.e., $V_g > -1.3$ V) by taking into account an extra attenuation of 7 dB resulting from the stray network. With higher resistance, the transmission becomes more sensitive to variations in capacitance. A careful analysis of the resonant frequency reveals that a shift from 786 MHz to 810 MHz occurs when V_g is varied from 0 V to -2.5 V. The total capacitance calculated from this frequency shift is shown in Fig. 2(b). The plateau around -1.5 V indicates complete depletion of 2DEG immediately under the gates.

In order to examine the sensitivity of this RF-PC electrometer, a sinusoidal signal at a frequency of $\xi = 1$ MHz and a power of -6.0 dBm was superimposed on a dc gate voltage of $V_g^0 = -1.405$ V. The input RF signal was fixed at $f_0 = 813$ MHz. In Fig. 3(a), two sidebands at 813 ± 1 MHz are shown in the output spectrum measured by the spectrum

analyzer with a bandwidth of 30 kHz. The sidebands come from the modulation of the PC's resistance ($R_q^0 \approx 42.4 \text{ k}\Omega$) by the ac gate voltage. Fig. 3(b) shows the measured and calculated sideband amplitudes, which decrease as the PC's resistance becomes larger. The oscillatory feature shown in the measured data is a result of the fluctuation in effective ac gate voltage caused by the changing stray capacitance as discussed earlier. As shown in Fig. 3(a), the noise-floor is about -102 dBm , corresponding to a noise voltage of $1.8 \text{ }\mu\text{V}$. The sidebands have an amplitude of about -90 dBm , i.e., the signal voltage is about $7 \text{ }\mu\text{V}$ and the signal-to-noise ratio (SNR) is 12 dB. The ac gate voltage can be estimated from the amplitude of the sidebands to be about $v_g \approx 63 \text{ mV}$. As an order-of-magnitude estimate, such an ac signal induces a charge variation of about $\Delta q = 0.707 C_g v_g \approx 140e$, assuming a gate capacitance of $C_g = 500 \text{ aF}$ [11]. The charge sensitivity of this RF-PC at V_g^0 can be estimated as $\delta q = \Delta q / (\sqrt{B} 10^{SNR/20}) \approx 2 \times 10^{-1} e / \sqrt{\text{Hz}}$ at $B = 30 \text{ kHz}$.

The present sensitivity is limited by the noise from within the spectrum analyzer, which is much larger than the intrinsic noises generated by the PC itself, such as shot noise and thermal noise [12]. The sensitivity can be significantly increased by using a cryogenic RF preamplifier, which has a low noise figure. Based on the present setup, an additional cold RF preamplifier with a noise temperature of 20 K and a gain of 20 dB will allow us to reduce the input RF signal from -40 dBm to -60 dBm . The shot noise is thus reduced by a factor of ten and the thermal noise at 4.2 K becomes the limiting factor. The overall charge sensitivity will be increased by a factor of 25. Furthermore, by carefully engineering the PC, the intrinsic sensitivity ($|dR_q/dV_g|$) could be increased by a factor of ten [8]. So, an overall charge sensitivity of $1 \times 10^{-3} e / \sqrt{\text{Hz}}$ at 30 kHz could be expected.

With a fixed design of the LC circuit and a given cryogenic RF preamplifier, the main way of improving the sensitivity is to engineer the PC so that both a larger ($|dR_q/dV_g|$) and a smaller R_q could be achieved at the chosen working point V_g^0 . As shown in Fig. 3(c), the benefit of a smaller resistance is a larger dT/dR_q . Furthermore, a smaller resistance means less thermal and shot noise. To have a large ($|dR_q/dV_g|$), one can set the working-point gate voltage in the middle of the sharp transition region between two quantized conductance plateaus. However, in this case, the dynamic range of the electrometer is strongly limited by the narrow transition range. A larger ($|dR_q/dV_g|$) can also be achieved by carefully engineering the design of PC even without conductance quantization. In this case, the RF-PC could provide both a high sensitivity and a large dynamic range. Comparing to resistive

RF-SETs, an RF-PC/RF-QPC made of the host semiconductor material is much easier to fabricate and integrate into the system under test. Furthermore, the quantum electron-waveguide nature of a QPC allows continuous weak quantum measurements [13, 14]. In addition to SETs and PCs, any controllable resistor/capacitor/inductor device could be integrated into this RF circuit as a fast and sensitive sensor. Some of the possible devices are MOSFET transistors, Zener diodes, and microelectromechanical capacitor sensors.

In conclusion, we have constructed and characterized an RF-PC with a charge sensitivity of $2 \times 10^{-1} \text{ e}/\sqrt{\text{Hz}}$ at 30 kHz which is limited by the noise from the RF measurement instrument at room temperature. By introducing a cryogenic RF preamplifier, lowering the input RF power, and improving the point-contact to have steeper pinch-off characteristic, the charge sensitivity could be increased to $1 \times 10^{-3} \text{ e}/\sqrt{\text{Hz}}$ or even higher. In addition to the fact that QPC electrometers promise an ultra-high sensitivity at quantum limit, we emphasize that it is easier to build an RF-PC than conventional RF-SETs. We also point out that any device, such as a variable resistor with a sharp control slope and a resistance below 100 k Ω , could be easily integrated into a similar RF circuit to form a both fast and sensitive detector/sensor.

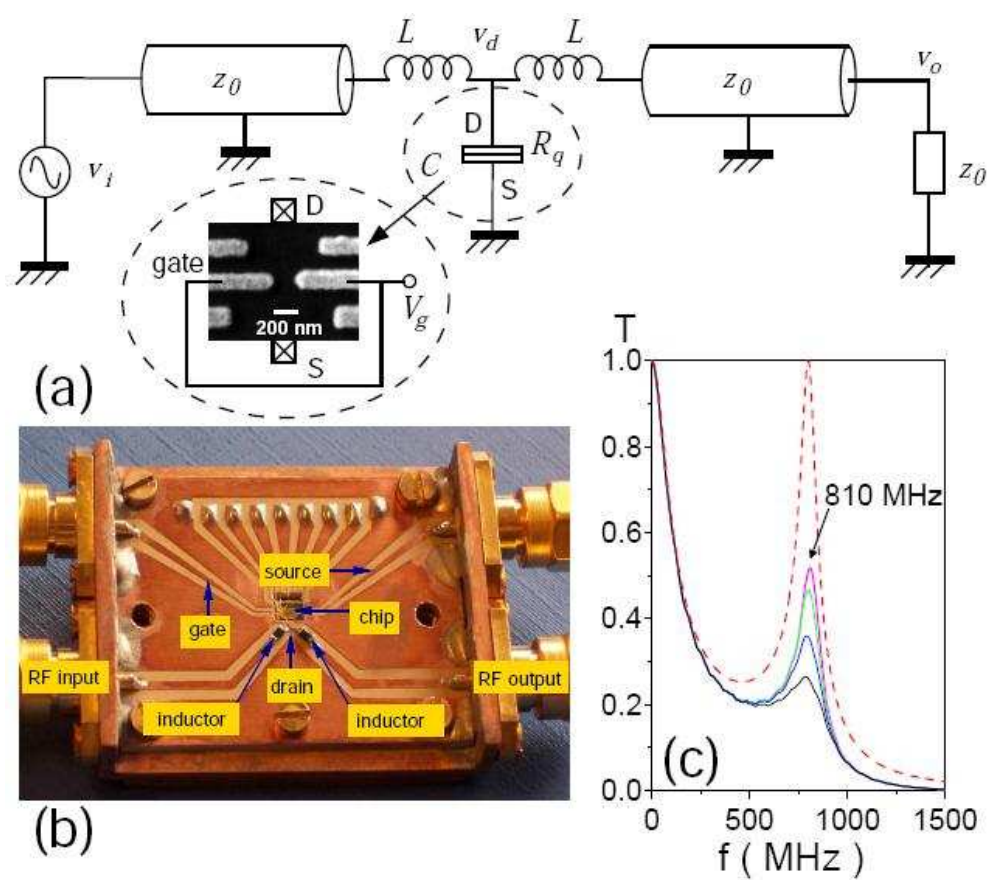
This work is supported by the U.K. Department of Trade and Industry and Hitachi Europe Limited under the Foresight LINK project “Nanoelectronics at the Quantum Edge”. The authors thank Dr. David G. Hasko for stimulating discussions and critical proof reading of the manuscript, and thank Dr. Richard J. Collier for the discussions on RF measurements.

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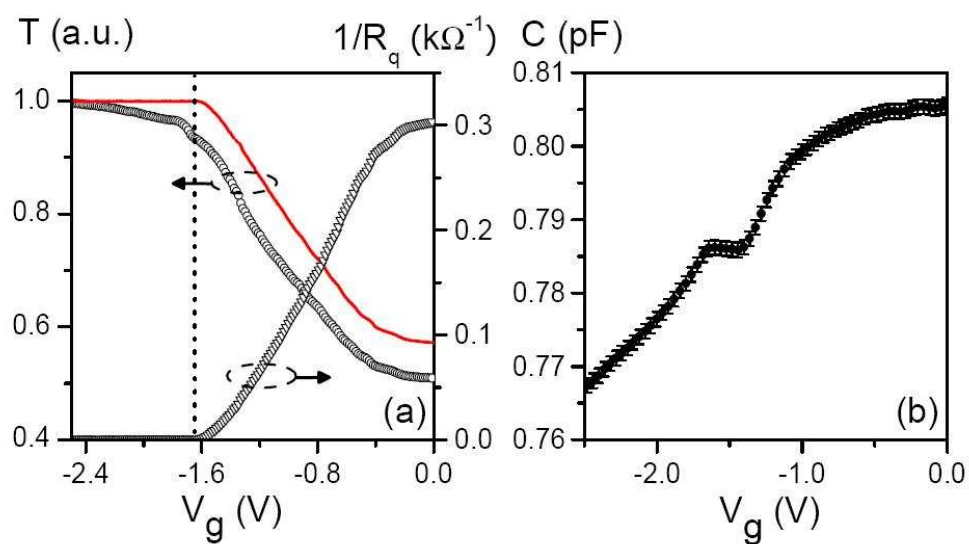
FIG. 1: (a) Circuit diagram for the RF-PC electrometer and a scanning-electron micrograph of the PC. (b) Realization of the RF-PC on a high-frequency printed-circuit board. (c) Measured transmission coefficient as a function of the frequency at 4.2 K (solid curves). From bottom to top, the dc gate voltage is 0 V, -1.0 V, -1.5 V, and -2.5 V, respectively. Calculated transmission coefficient is shown as the dashed curve.

FIG. 2: (a) The correspondence between the resonant transmission coefficient and the PC's resistance. The vertical dotted line marks the pinch-off voltage (-1.65 V). The solid curve represents the calculated resonant transmission amplitude. (b) The total capacitance of the PC is varied by the gate voltage.

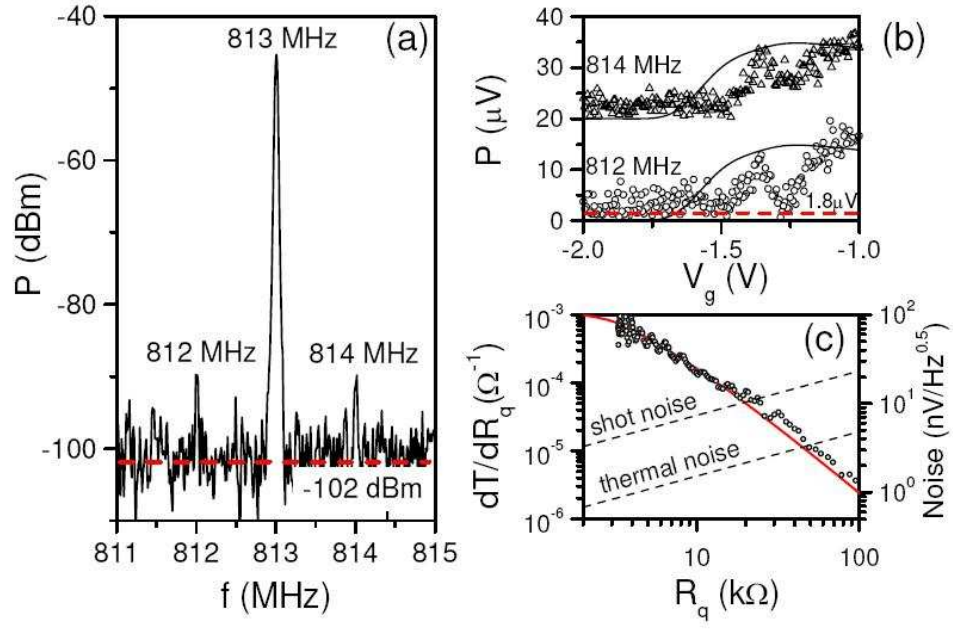
FIG. 3: (a) Transmission spectrum measured by the spectrum analyzer. (b) Linear-scale amplitude of the sidebands as a function of the gate voltage compared with the simulations (solid curves). For clarity, the data for sideband at 814 MHz are shifted upward by $20 \mu\text{V}$. (c) Measured $dT/dR_q - R_q$ curve is compared with the simulation (solid curve). The dashed lines are calculated shot noise and thermal noise across the PC. The PC has a resistance of $42.4 \text{ k}\Omega$ and is biased with a constant voltage $v_{ds}^{rms} \approx 17 \text{ mV}$ at 4.2 K.



Qin *et al.*: Figure 1/3



Qin *et al.*: Figure 2/3



Qin *et al.*: Figure 3/3